TECHNICAL INFORMATION

The Technical Information Section consists of a collection of terms, technical notes and measurement techniques frequently utilized to describe solid state microwave amplifier performance. Information is presented alphabetically according to the following parameter list:

For additional information and assistance in describing or specifying amplifier performance, please contact the Microwave dB Applications Engineering staff.

Bandwidth

The bandwidth of an amplifier, as defined by the manufacturer, is the frequency range over which performance parameters are specified. It should be noted that the operational bandwidth of a microwave amplifier may be much greater than the specified performance bandwidth.

Burnout, RF

RF Burnout is the permanent degradation and/or failure of an amplifier due to the application of a pulsed and/or CW RF input signal with sufficient power to exceed the amplifier's safe operating level.

Safe Operating Level:

Without protection circuitry, the input level at which amplifier performance will degrade is, in general, a function of the geometry and the physics of the transistor used in the first amplifier stage.

The initial effect of application of excessive input power to a bipolar amplifier is a slight but permanent increase in the amplifier noise figure. As the input level is further increased, the amplifier noise figure will continue to degrade and a point will be reached at which the amplifier gain will suffer permanent degradation. Of course, the ultimate effect is catastrophic failure of the amplifier. GaAs FET amplifiers typically exhibit a catastrophic failure without being preceded by degradation.

Unless otherwise specified, all bipolar amplifiers manufactured by Microwave dB are guaranteed to withstand continuous application of at least +13 dBm CW input power for an indefinite period without degradation. GaAs FET amplifiers will withstand +20 to +23 dBm depending on whether they are of balanced or single ended design. However, it should be noted that many amplifiers in this catalog are capable of withstanding input power up to several watts (without additional protection circuitry) depending on the duration of the input signal as well as the particular unit selected.

Specifications regarding input power-handling capability must address the duration and amplitude of the peak power as well as CW or average power of the input signal.

Input Protection:

When an amplifier is not capable of surviving a specific power level, Microwave dB can provide circuitry that will protect the unit from the higher power levels, which cause burnout. These circuits consist of PIN (sometimes in conjunction with Schottky) diodes that are integrated within the amplifier. Typical input protectors will allow the amplifier to withstand peak power as high as 100 watts for up to 1 microsecond and CW or average power up to 5 watts. Typical insertion loss for this type of limiter is 0.4 dB.

Input protection circuits are usually reflective and the maximum input match should be defined for both linear input signal levels and limiting input signal levels. The Microwave dB Applications Engineering Staff should be consulted regarding any particular input protection requirements.

Compression Point

The compression point of an amplifier is traditionally defined as the output power level at which the gain of the amplifier decreases from its linear (small signal) level by 1dB. This point also frequently defines the upper limit of the amplifier's dynamic range (see `Dynamic Range`). A general discussion of compression effects is given under `Power Output`.

Dynamic Range

The Dynamic range of an amplifier is the range or input power over which the amplifier will linearly reproduce the input signal at the output. The lower level is set by the minimum detectable signal and is determined by the amplifier's noise figure and system bandwidth. The upper level, depending on the specific system application, is usually either determined by the power level at which the amplifier begins to compress (usually the 1 dB compression point) or the acceptable distortion (intermodulation or harmonic) level.

Gain

As used by an amplifier manufacturer, gain is defined at the transducer power gain expressed in dB when the source and load are matched, thus:

$$\text{Transducer Power gain} = 10 \text{ Log } \frac{\text{Power delivered to the load}}{\text{Power available from the source}}$$

The voltage gain of the amplifier is therefore the square root of the power gain when source and load impedance are equal.

Unless otherwise stated, any gain discussions in this text assume:

- The gain reference system has a 50 ohm characteristic impedance, and
- Gain is measured in the linear (small signal) region of the amplifier before compression begins to take effect (see figure 8 under `Power Output`).

Gain Realization:

Microwave dB technology is based on the use of bipolar and GaAs FET microwave transistor technologies. Typically, bipolar devices are used at frequencies as high as 5 GHz. GaAs FET transistors are used in designs that cover the entire frequency spectrum from several MHz to (presently) 40GHz.

Each amplifier stage contains all the circuitry required to realize a complete amplifier. Stage gains vary from as high as 20 dB at low microwave frequencies to 4dB for very high frequencies or very large bandwidths, or in some high power requirements. Overall amplifier gain requirements are achieved by cascading sufficient stages.

Each stage may be either "single ended" or "balanced". A single ended stage is one that utilizes one microwave transistor to produce, when practical, a satisfactory combination of VSWR, gain, flatness, noise figure and output power. A balanced stage consists of two single ended stages driven in parallel by the outputs of a quadrature hybrid. The two amplified outputs are then recombined in a second quadrature hybrid to give a single output signal. The major advantage of the balanced amplifier is that input and interstage VSWR is well controlled allowing very broadband amplifiers to be constructed from unit stage building blocks (referred to as gain modules). Other advantages of the balanced design are a nominal 3 dB increase in power output and intercept point and a "soft fail" mode such that amplifier gain will drop approximately 6 dB if one device in a balanced design fails, as opposed to catastrophic failure in a single ended design. The disadvantages of the balanced approach are the increased cost, size and power requirement of an additional transistor and supporting circuitry for each amplifier stage. These costs are usually more than offset when compared with alternative methods that may be used to produce similar results.

Gain Variation/Gain Window:

Gain Variation refers to the total peak-to-peak gain change over all specified operating conditions. Gain variation with frequency, with all other parameters held constant, is referred to as gain flatness. Gain variation with temperature, with all other parameters held constant, is referred to as the temperature stability of gain. Gain variation with bias, with all other parameters held constant, is referred to as the time stability of gain. The total gain variation of the amplifier may be written:

$$\Delta G = 2(\pm \Delta G_f \pm \Delta G_T \pm \Delta G_b \pm \Delta G_t)$$

where:

ΔG = the total gain variation

ΔG_f = the gain flatness
ΔG_T = the gain variation with temperature
ΔG_b = the gain variation with bias
ΔG_t = the gain variation with time

Here it should be noted that the terms G_i have been defined with a ± term indicating that they are each equal in magnitude to ½ their peak-to-peak change. They may also behave at certain values of their variable to offset each other. ΔG may also be defined as the gain window.

Microwave dB amplifiers display minimal effects of bias and time stability. All standard MIC amplifier designs contain a temperature stable internal voltage regulator which allows the user to vary input voltage supply by at least ±10% with no measurable associated gain shift. Most of Microwave dB's discrete amplifiers are specified at ±1% maximum bias voltage change.
If power supplies change more than these specified amounts, the user should so state and tighter regulation may be incorporated.

Gain stability with time can be influenced by changes inherent in the semiconductor devices which may occur over short or long intervals of time. Early microwave transistor amplifiers were subject to shifts that occurred with time, but present microwave device materials and processing technologies have made such changes all but disappear. Another stability effect, sometimes confused with time stability, is actually a temperature effect and is due to self-heating of the amplifier at turn-on. This effect is more pronounced in power designs where internally generated heat may cause a gain reduction to occur during the first 5-10 minutes after turn-on until the amplifier has stabilized in its ambient environment.

Gain Flatness / Ripple / Slope:

Gain Flatness over the amplifier bandwidth is defined at a constant temperature (nominally +25°C unless specified otherwise). Although the gain shape (see Figure 1) will be similar at other values of temperature, minor changes in flatness may occur over the operating temperature range. Fine grain gain change over a small segment of frequency within the amplifier bandwidth is referred to as ripple. Gain slope is the incremental change in gain with respect to frequency at a specific frequency. It should be thus noted that the shape of the gain response is not controlled through a gain flatness specification.

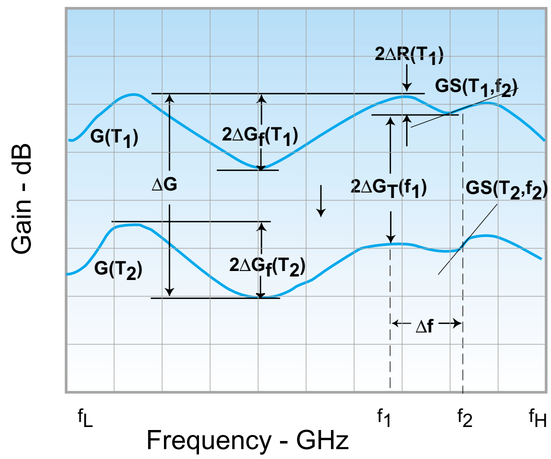

Figure 1: Gain Shape

$f_H - f_L$	=	amplifier bandwidth (specified)
$G(T_i)$	=	typical gain shape of amplifier at temperature T_i
$\Delta G_f(T_i)$	=	flatness of amplifier at temperature T_i
$\Delta R(T_i, \Delta f)$	=	ripple over frequency range Δf at temperature T_i
$GS(T_i, f_2)$	=	gain slope at frequency f_2 and temperature T_i
$\Delta GT(f_1)$	=	stability of gain with respect to temperature at frequency f_1, over temperature range $T_2 - T_1$
ΔG	=	gain window over temperature T_1 to T_2 and frequency f_L to f_H

Typically gain flatness becomes worse as the number of amplifier stages increases and/or as the amplifier bandwidth increases. This is due principally to VSWR interaction between the output of one stage and the input of the following stage. Over broad bandwidths, the phasing of these mismatches will go through maxima and minima creating a characteristic ripple in the gain response. Although the magnitude of these excursions can be reduced over a finite bandwidth by tuning the amplifier, it is sometimes accomplished at the expense of other parameters. This may be especially true in low gain (1-2 stage) amplifiers where good flatness can have a significant effect on noise figure, power output, or input and output VSWR if these latter parameters are critical.

Temperature Stability of Gain:

The physics of both microwave bipolar transistors and GaAs FET's causes an inherent gain reduction as the amplifier temperature is increased. This gain change is dependent on the devices used, the operating frequency, and the actual operating temperature. A typical gain versus temperature coefficient range per amplifier stage is:

0.0009 to 0.015 dB/$_{\text{degree C}}$

Thus, a four-stage X-Band design over a temperature range of –54 to +85°C should vary in gain somewhere between 5.00 and 8.34 dB depending on the devices employed. Figure 2 shows the measured variation of gain with temperature for a 6 to 12 GHz amplifier. The variation of gain over the temperature range is slightly frequency dependent and varies from 5.1 to 6.3 dB.

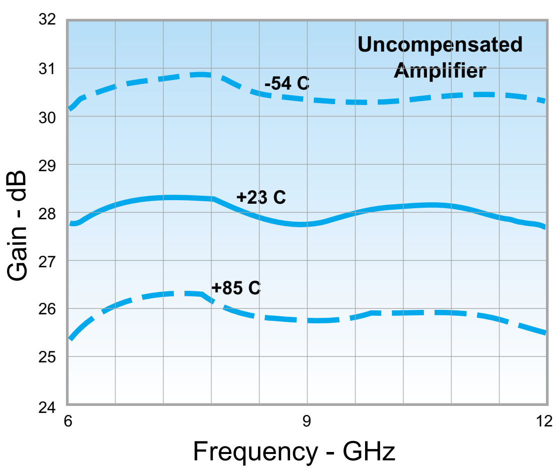

Figure 2:
Uncompensated Gain - 6-12 GHz Amplifier

Temperature Compensation of Gain:

Amplifier temperature compensation is accomplished by varying some voltage sensitive gain element in the amplifier, typically by means of a thermistor network. For bipolar designs, this may be accomplished by simply varying the stage voltage over the temperature range, if the amplifier bandwidth, temperature range, and/or degree of compensation required is not extreme. For most GaAs FET amplifiers and critical bipolar designs, a voltage variable PIN diode attenuator is incorporated in the amplifier and its attenuation level is changed over the operating temperature range to compensate for the amplifier variation. Typical compensation networks reduce the variation over temperature by 50% to 70%.
It should be noted that such an attenuator network will have the net effect of reducing overall amplifier gain since the reference point will be the high temperature gain of the amplifier and the "off" attenuator insertion loss will be 0.5 to 2.0 dB. Moreover, in integrated circuit designs, the addition of an attenuator module stage will increase the amplifier length by an additional stage. Figure 3 below shows the effect of temperature compensating the 6 to 12 GHz amplifier with a PIN attenuator module. The gain window over the temperature range of –54 to+85°C has been reduced from 6.3 dB to 2.3 dB.

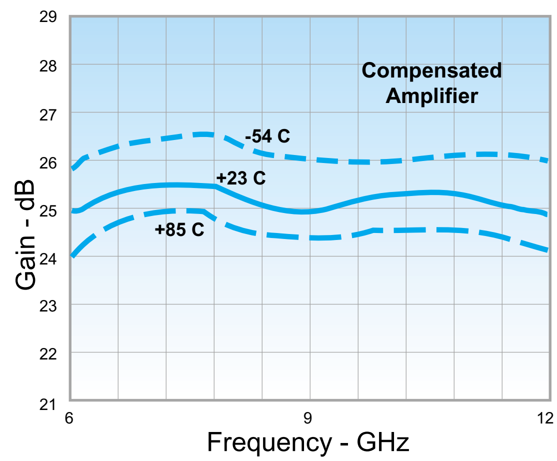

Figure 3: Compensated Gain Model

Gain Measurement:

Two basic problems occur when making precise gain measurements:

- Referencing the gain to an accurate standard
- Eliminating the effects of mismatch on gain ripple

Figure 4 shows a precision gain measurement test set. By using the frequency response test set in a ratio mode, the effects of power variation from the sweep generator are cancelled. In a typical measurement, the output detector is placed on the test arm of the power splitter and a reference is obtained. The amplifier and calibrated attenuator are then connected and the frequency response test set is used to determine the gain deviation of the amplifier from the reference attenuator. Errors in the gain measurement arise from:

- Calibration errors in the test equipment
- Uncertainties in the absolute value of the reference attenuator; typically ±0.1 dB/10 dB

of attenuation
- Effects of test mismatch

The latter effect may be illustrated by analyzing the interface VSWR's, V_i, as shown in Figure 4 below.

When point A is connected to point B to set a reference level, the reference uncertainty is:

$$U_r = \pm 10 \text{ Log} \left(\frac{V_1V_6 + 1}{V_1 + V_6} \right) \text{ in dB}$$

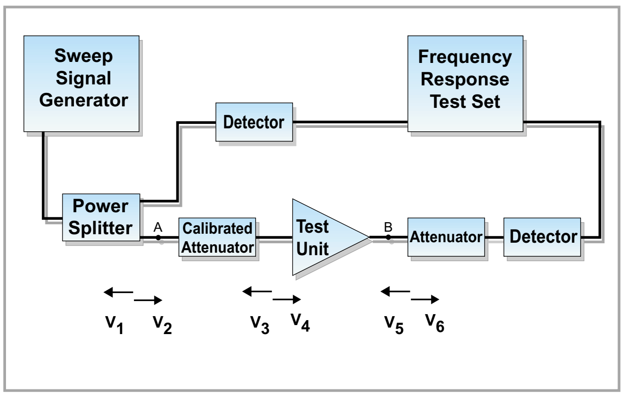

Figure 4: Gain Measurement Test Set

When the amplifier and the calibrated pad are inserted in the setup, additional interface VSWR's are generated. It can be shown that the total uncertainty, including the reference VSWR uncertainty, is given by:

$$U_V = \pm 10 \log [A \cdot B \cdot C \cdot D]$$

Where:

$$A = \left(\frac{V_1V_2 + 1}{V_1 + V_2} \right) \qquad B = \left(\frac{V_3V_4 + 1}{V_3 + V_4} \right)$$

$$C = \left(\frac{V_5V_6 + 1}{V_5 + V_6} \right) \qquad D = \left(\frac{V_1V_6 + 1}{V_1 + V_6} \right)$$

For a typical measurement setup, the equipment VSWR's V_1,V_2,V_3,V_6, are less than 1.1:1. Using 1.1:1 for these quantities, and assuming the amplifier input and output VSWR's V_4 and V_5 are 1.5:1, the resulting gain measurement uncertainty is ±0.205 dB. If the amplifier VSWR's are 2.0:1 the measurement uncertainty rises to ±0.315 dB. It should be noted that if all equipment VSWR's were 1.0:1 there is no gain measurement uncertainty due to amplifier VSWR.
The reference attenuation uncertainty must be added to the above quantity to arrive at the total measurement uncertainty.

In making gain measurements, a number of other precautions are important. Care must be exercised to ensure that the harmonic content of the sweep generator is low. The broadband detectors cause the harmonic power to affect the display when obtaining a reference. An amplifier which doesn't amplify the harmonic will appear to have lower than actual gain. It is also important to prevent inadvertent increases in amplifier input power, as may sometimes happen when increased measurement dynamic range is required. This can easily drive the amplifier into compression which will cause the measured gain to be low. Gain measurements should be performed at output levels at least 15 dB below the 1 dB gain compression point of the amplifier. The proper choice of the calibrated reference attenuator will assure that the amplifier is not over driven.

Isolation

An amplifier's (reverse) isolation is the power delivered to a matched load connected to the input of the amplifier when the output is driven from a unity power source of the same characteristic impedance (typically 50 ohms). Thus, isolation is simply the reverse gain of the amplifier (see Gain measurement). Generally, the reverse isolation (attenuation) of the amplifier is on the order of 1.5 to 2.0 times the forward gain (e.g., a 20dB gain amplifier would be expected to have 30-40 dB of isolation). If the gain exceeds the isolation at some frequency, the amplifier would be unstable at that frequency and oscillate. Even if the isolation is greater than but approaches the gain, a significant ripple component would result. Care is exercised in amplifier design to assure that such a condition does not exist.

Matching/Tracking

These parameters measure the extent to which individual amplifiers of a particular type, on a production lot basis or more simply as pairs, mirror each other with respect to one or more performance characteristics over the frequency range. Figure 5 shows a typical frequency characteristic of a pair of amplifiers.

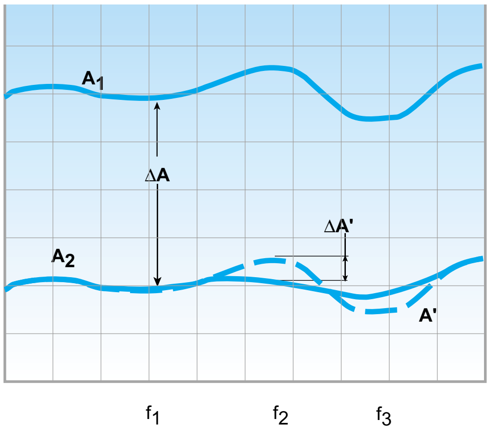

Figure 5: Matching/Tracking

The matching of A_1 to A_2 at frequency f_1 is its offset ΔA. The tracking of A_1 to A_2 at frequency f_1 is the difference between the two curves if the average offset over bandwidth is subtracted out. A'_1 is the overlay of the A_1 characteristic on the A_2 characteristic after the average offset has been removed. A maximum difference of $\pm\Delta A'$ occurs at frequencies f_2 and f_3. The tracking window is $2\Delta A'$ or characteristic A_2 tracks A_1 within $\pm\Delta A'$.

Tracking thus implies that the user has some means to remove the average offset between amplifiers. For gain, this is accomplished via the incorporation of an attenuator or a frequency sensitive equalizer. For phase, a fixed line length would be added. For the amplifier manufacturer, matching is often achieved at the expense of tracking. For example, to achieve gain matching of two units, one unit may have to be deliberately detuned for gain which may cause additional ripple in its gain characteristics. Matching is more difficult for the amplifier manufacturer to achieve than tracking. Phase is more difficult to match than gain. The tightness of matching and/or tracking specifications impact cost, delivery times, and in some cases amplifier size. Microwave dB Applications Engineering should be contacted for assistance in specifying matching/tracking parameters.

Noise

Noise is present in all electromagnetic media at temperatures above absolute zero. Being random in nature it will obscure electronic signals if the signal power is less than the noise power present. Noise is induced in systems due to its reception by the system's antenna and/or its generation by the "non-zero" temperature of system components.
A microwave amplifier will amplify all noise within its operating bandwidth as well as generate additional internal noise in the amplification process.

Noise Figure:

The magnitude of internally generated noise is typically quantified by the amplifier's Noise Figure F, defined as the input signal to noise ratio divided by the output signal to noise ratio.

Since the output noise power is basically the amplified input noise plus the noise generated by the amplifier, the noise figure is a measure of internally generated noise, and is related to the amplifier's noise temperature T_e (°K) by the equation:

$$F = 10 \text{ Log} \left(1 + \frac{T_e}{290} \right)$$

The noise figure for microwave amplifiers varies from below 1 dB at low frequencies to 10 dB or more at very high frequencies. In addition to its frequency dependence, it is a function of the transistor used and the design of the circuit including both bias and impedance matching considerations. In all applications where noise figure is of prime concern, care should be exercised to minimize circuit losses in front of the amplifier. (e.g. limiters and/or filter) since their insertion loss adds directly to the noise figure.

Temperature Variation of Noise Figure:

Both GaAs FET and bipolar transistor amplifiers have noise characteristics that increase with increasing temperature. This relationship can be approximated by the following empirically derived equation:

$$F(T) = 10 \text{ Log} \{ (N - 1) \tau^a + 1 \}$$

Where:

$$N = \text{Log}^{-1}\left(\frac{F_o}{10}\right)$$

F_O = noise figure at room temperature, T_o°K

$$\tau = \frac{T}{T_O}$$

T = physical amplifier temperature °K

a = experimentally derived constant typically between 0.5 and 1.5 for bipolar transistors and approximately equal to 1 for most GaAs FET transistors

The above equation will provide reasonable results for bipolar transistors above approximately -40°C (233°K). For GaAs FET's, however, the equation has been found to hold down to cryogenic temperatures. Unlike bipolar transistors which cease to function below about -80°C, GaAs FET's will continue to operate at temperatures near absolute zero. In fact, the noise figure will continue to decrease with decreasing temperature down to about 40°K. For example, Microwave dB has performed experiments on GaAs FET amplifiers at very low temperatures wherein the unit was immersed in liquid nitrogen (77°K) and noise measurements performed. The noise figure of a 4 GHz amplifier module was reduced from its room temperature value of 1.5 dB (Te = 120°K) to 0.51 dB (Te = 36°K).

Noise Figure Measurement:

In general the typical methods used to measure noise figure are all based on the technique described in figure 6.

With the amplifier input connected to respective terminations at temperature T_H and T_C, the power measured at the receiver output is:

$P_H = k(T_e + T_H)\,B\,G$ (when connected to a hot load)
$P_C = k(T_e + T_C)\,B\,G$ (when connected to a cold load)

Where:

k = Boltzman's constant = 1.38×10^{-23} watts-seconds/°K
T_e = Equivalent noise temperature of the amplifier
B = Receiver bandwidth (Hz)
G = Amplifier and receiver gain (numeric ratio)

The ratio of these two measurements is known as the "y" factor.

$$y = \frac{T_e + T_H}{T_e + T_C} \quad \text{or} \quad Te = \frac{T_H - yT_C}{y-1}$$

And

$$\text{Noise Figure, } F = 10\,\text{Log}\left\{1 + \frac{T_e}{290}\right\} \text{ in dB}$$

From the above equations, it is obvious that the accuracy to which T_H and T_C is known, determines the accuracy of the noise figure measurement.

The traditional method that is used to produce a termination at a precisely known temperature is to immerse a resistive load in liquid nitrogen. Since the boiling point of nitrogen is known with great accuracy, so is the temperature of the load. The typical measurement procedure is to connect the "cold" load to the amplifier, note the output and then connect a room temperature (or hotter) load of known physical temperature. The variable attenuator is then adjusted to produce an output level equal to that when the cold load was connected. The y factor is therefore equal to the attenuation difference between the two measurements.

The above procedure is tedious, time consuming and does not lend itself to a production test environment. However, if requested Microwave dB can utilize this method to perform critical measurements on very low noise amplifiers for an additional charge. Microwave dB also uses the foregoing method to calibrate laboratory standard amplifiers to be used as production standards. The calibrated amplifiers allow the use of automatic noise figure measuring equipment that will result in measurement accuracy's very nearly equal to that provided by a cryogenic method.

Automatic noise measuring equipment operates on basically the same principle. However, instead of manually switching loads, the "Automatic Noise Figure

Meter" (ANF) utilizes a "noise diode" that is alternately switched on and off at a kilohertz rate. When the diode is on, it produces the effect of a very noisy or high temperature load; when off it is essentially at room temperature. The ANF performs a calculation of noise figure and displays the result on a digital display. The accuracy of the measurement depends on the precision to which the "excess noise ratio" of the diode noise source is known as well as the accuracy of the equipment used. This method by itself usually results in noise figure uncertainties on the order of 0.2 to 0.4 dB.

Through the use of the aforementioned standard amplifiers to calibrate the entire system, noise figure measurement uncertainties can be reduced to less than 0.05 dB in a production environment.

Figure 6: Noise Figure Measurement Set

Phase Distortion

Phase distortion may be defined as the degree to which the phase characteristics of an amplifier differs from the propagation of electromagnetic energy through an ideal transmission media.

Phase Linearity:

Phase linearity is frequently employed to define the phase distortion introduced by a solid state amplifier. The deviation of phase shift from the ideal straight line is defined to be the phase linearity of an amplifier. Figure 7A illustrates a phase shift versus frequency plot. The slope of the best fitted straight line through the phase shift versus frequency plot is known as the phase delay and determines the steady state phase relationship between input and output signal at a fixed frequency.

$$\text{Phase delay} = T_p = -\frac{F}{\omega}$$

Phase delay is sometimes referred to as carrier delay.

As figure 7A indicates, the phase distortion may be defined in terms of phase linearity; $\Delta F(\omega)$ where,

$$\Delta F(\omega) = F(\omega) - (-T_p\,\omega)$$

This figure graphically illustrates a phase linearity of ±A radians maximum. Phase linearity is sometimes referred to as ripple; however, phase linearity is the preferred term to define the phase distortion illustrated in figure 7A. Phase ripple is more frequently used to define a component of group delay distortion.

Group Delay Distortion:

In certain applications, it is more convenient and appropriate to discuss phase distortion in terms of group delay. Group delay, Tg (ω), is defined to be the negative derivative of the phase shift characteristic and determines the propagation delay of signal energy or information. This delay is often referred to as "envelope delay". Group delay and group delay distortion definitions may be derived from the previous phase linearity equation to be:

group delay,

$$T_g(\omega) = -\frac{dF(\omega)}{d\omega} = T_p + -\frac{d(\Delta F\,\omega)}{d\omega}$$

group delay distortion, $\Delta F(\omega) = T_g(\omega) - T_p$

Hence, phase distortion may also be defined in terms of group delay distortion, $\Delta F(\omega)$, which is the deviation of group delay, $T_g(\omega)$from a constant group delay, T_p. As this deviation goes to zero, the group delay becomes equal to the phase delay as would be expected for an ideal network. Many amplifier networks exhibit complex group delay functions such as in figure 7B. In certain communications and radar applications, these complex group delay functions are approximated over limited band segments using the following quadratic equation:

$$\Delta F\ (\Delta f) = \frac{\Delta F\ (\Delta\omega)}{2\pi} = \pm\,\tau_R + \tau_L f + \tau_p f^2 \quad \text{(in nsec)}$$

Where

R = ripple component in nsec (peak to peak)

L = Linear component in $\frac{\text{nsec}}{\text{MHz}}$

P = parabolic component in $\frac{\text{nsec}}{\text{MHz}^2}$

Linear, parabolic and ripple components group delay are more frequently specified in communications applications. When specifying phase distortion parameters, it is advisable to contact the Microwave dB Application Engineering staff. They will be pleased to assist in establishing cost effective specifications and test methods.

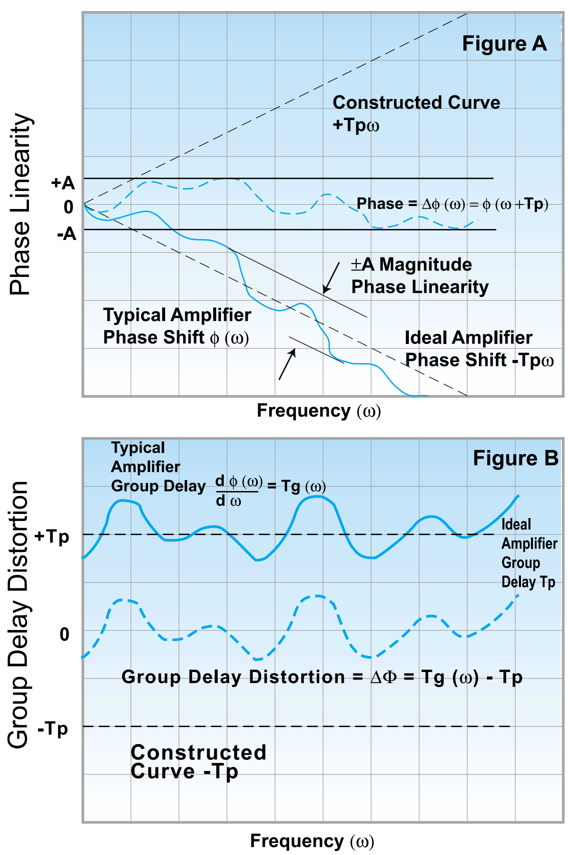

Figure 7: Phase Distortion Characteristics

Power Output

The output power capability of an amplifier is usually determined by specifying either the 1 dB compression point or the saturation level.

Depending on the design, the saturated output power capability will generally exceed the 1 dB compression point by 1 to 5 dB.

1 dB Compression and Saturation:

Over the linear dynamic range, the output power will exceed the input power by the amplifier's gain. As the input power is increased and the amplifier begins to saturate, an output level will be reached at which the gain is reduced by 1 dB. This level is defined as the output power at 1 dB compression and is typically used to specify an amplifier's output power capability.

As the input power is further increased, the amplifier output power will reach a point where it is relatively constant as a function of input power. This output level is the saturated output power of the amplifier.

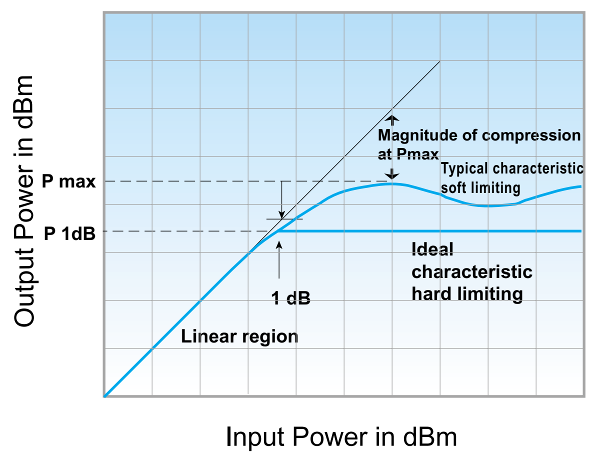

Figure 8: Power and Limiting Characteristics

Output Power and Limiting Characteristics:

Small signal amplifiers are normally designed for minimum noise and/or maximum gain per stage. These designs typically provide a greater output power difference between their 1 dB compression point and maximum output power than power or limiting amplifier designs. As a rule of thumb, low noise amplifiers will be less than 0.1 dB compressed for power output levels up to 10 dB below the 1 dB compression point and less than 0.4 dB compressed at power levels up to 6 dB below the 1 dB compression point. Maximum power output will be 3 to 5 dB greater than the power output at the 1 dB compression point. Small signal amplifiers have nominal power outputs at the 1 dB compression point of up to +20 dBm.

Power amplifier designs, unlike small signal designs, are not impedance matched to produce maximum gain, but are designed to produce maximum power and efficiency. Also, power design may have several stages that compress simultaneously as opposed to low noise units where the output stage usually is in compression first. As a consequence, the maximum power output of a power amplifier may be only 1-3 dB above the power output at the 1 dB compression point. Typical microwave transistor "power" amplifiers have power output ranging from +20 to +33 dBm at the 1 dB compression point.

Limiting amplifiers are used in applications such as direction finding (DF) and instantaneous frequency measurement (IFM) systems. Limiting amplifiers may be segregated into two categories: "soft limiting": amplifiers and "hard limiting" amplifiers. "Soft limiting" amplifiers gradually saturate until input power increases have little effect on the output power level. "Hard limiting" amplifiers abruptly change from the linear to the limiting region. A major design consideration is how abrupt the transfer curve changes from the linear to the limiting region. "Hard limiting" amplifiers usually require the difference between 1 dB compression and saturation power output to be less than 3 dB.

"Soft limiting" amplifiers are not as concerned with output power difference since they are primarily used in the limiting region. A "soft limiting" amplifier is designed with sufficient gain to drive the amplifier well into compression for the minimum input signal level. Obviously, the broader the input signal range the more difficult the "soft limiting" amplifier design. Figure 8 illustrates the difference between soft and ideal limiting characteristics. It should be noted that recent results with dual gate FET's indicate promise in certain "hard limiting" applications and many "soft limiting" designs are replacing TWTs with better performance, lower intermods and longer operating life.

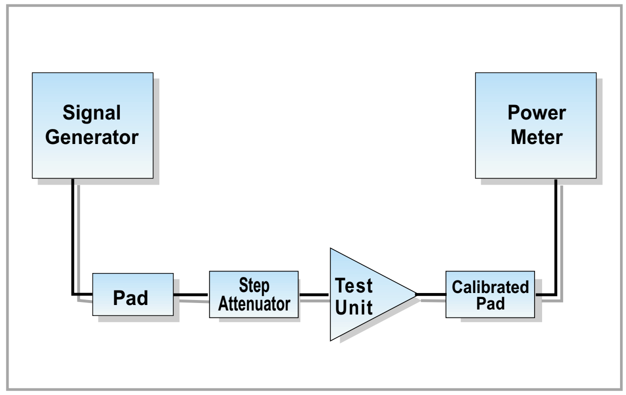

Figure 9: Output Power Test Set

Output Power Measurement:

Figure 9 shows a typical setup for power output measurement. In order to measure output power at 1 dB compression, the input power is increased in steps until the change in output power is only 9 dB for a 10 dB change in input power.

Conversely, the output power may be set to the specified 1 dB compression point. If the input power is then reduced by precisely 10 dB, the compression can be measured as the difference between 10 dB and the actual reduction in output power. For example, if the output power reduction was 9.3 dB when the input changed 10 dB, the amplifier would be 0.7 dB compressed at that output level.

Because of the simplicity of the latter method, unless otherwise required, Microwave dB provides data showing the degree to which the amplifier is compressed at the specified 1 dB compression point.

Recovery Time

Recovery time is defined as the time required for an amplifier to recover from a saturated condition to within 1 dB (unless otherwise specified) of its small signal gain.

Circuit Considerations:

The recovery time of an amplifier is a complex function of the biasing and decoupling networks used in the design of the unit. Designs that provide excellent stability of transistor bias with respect to device to device variation, temperature and time will frequently result in relatively slow recovery characteristics. Other design considerations that affect recovery time include techniques that prevent low frequency oscillations during saturation and provide isolation of the amplifier gain stages from power supply induced noise and RFI. Additionally, amplifiers that are designed to enhance recovery time with minimal sacrifice in the aforementioned parameters will usually be less efficient with respect to power supply requirements (this is particularly true of low frequency bipolar transistor amplifiers).

The amplifiers shown in this catalog have been designed to provide the best trade off between recovery time and other circuit consideration mentioned above. Recovery time of standard products will range from several microseconds for low frequency bipolar amplifiers to several nanoseconds for high frequency MIC GaAs FET amplifiers. However, as an option (usually for additional cost), all Microwave dB amplifiers can be constructed using alternative design techniques that will enhance recovery time with minimal sacrifice in other parameters.

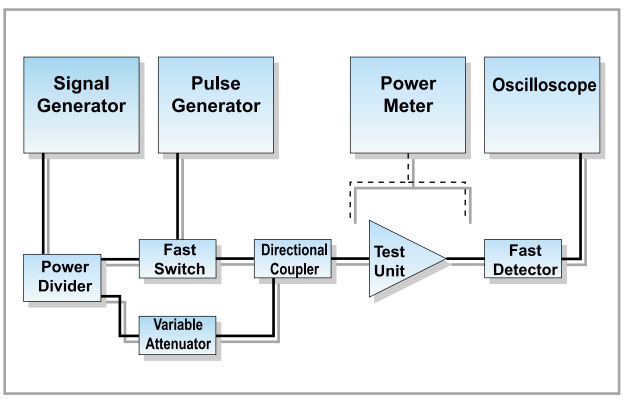

Figure 10: Recovery Time Test Set

Recovery Time Measurement:

In Figure 10 the signal generator and variable attenuator are adjusted such that with the switch on, the input power is sufficient to saturate the amplifier at a specified level and with the switch off the CW output power is at least 5 dB below the 1 dB compression point.

The resulting input and output wave form displayed on the oscilloscope will be shown in Figures 11A and 11B respectively. Recovery time can be measured directly from the display.

Spurious Signals

Spurious response from an amplifier is defined as any signals appearing at the output that were not part of the input wave form. The three most important forms of spurious response are intermodulation distortion, harmonic distortion and non-harmonic spurious response.

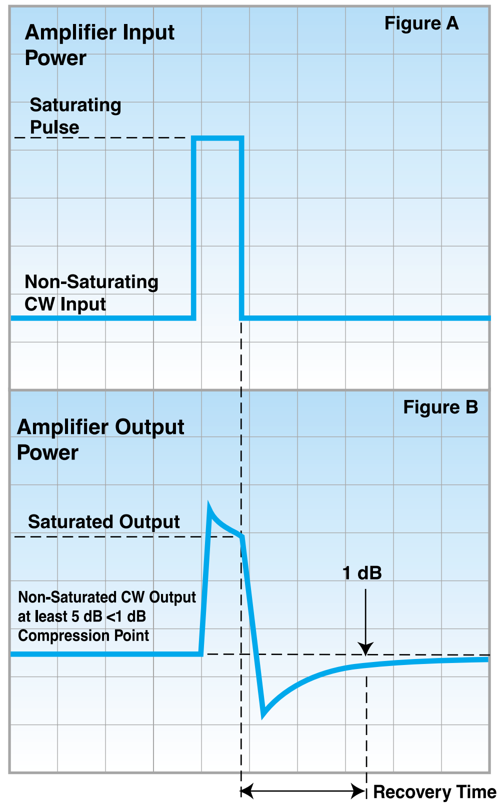

Figure 11: Recovery Time Wave Forms

Intermodulation Distortion (IMD):

IMD products result from the mixing of two or more input signals of different frequencies. The mixing is a consequence of the non-linearity of the amplifier gain as a function of input power.

To illustrate this, consider an amplifier with signals at fundamental frequencies f_1 and f_2 at the input. The output will contain signals at the following frequencies: $nf_1 + mf_2$ where n, m = 0,±1,±2,±3….The order of the IMD product is defined as:

$$i = |n| + |m|$$

With respect to each IMD product, there is an imaginary power level, P_i, which is defined as the Intercept Point (IP) of order i. This term is frequently used to specify the IMD performance of an amplifier.

The intercept point has become an industry standard used in specifying the intermodulation distortion of a microwave amplifier (or any non-linear microwave component). This parameter is usually measured by applying two signals to the input of an amplifier such that the output signals are equal. The intercept point is then calculated from:

$$IP_i = P_0 + \frac{(P_0 - P_i)}{(i-1)} \quad \text{in dBm}$$

Where:

IP_i = Intercept point of order i (dBm)

P_o = Output power of each of the two test frequencies (dBm)

i = Order of IMD products (e.g. second, third etc.)

P_i = Output power of the IMD product of order i (dBm)

Conversely, the IMD products are usually specified for equal output signals and can be easily calculated from the above equation. However, the question of unequal IMD products resulting from unequal output signals frequently arises. This question can be answered from the following equation which is a more general form of that given previously.

$$P(n,m) = IP_i - |n| D_1 - |m| D_2 \quad \text{in dBm}$$

Where:

$P(n,m)$ = Power level (dBm) of the IMD product at frequency

f = $nf_1 + mf_2$

IP_i = Previously defined

D_1 = $IP_i - P_1$ (dB)

D_2 = $IP_i - P_2$ (dB)

P_1 = Power output at signal frequency f_1 (dBm)

P_2 = Power output at signal frequency f_2 (dBm)

The terms of this equation are illustrated in Figure 12 for the typical case of third order IMD resulting from unequal output powers at frequencies f_1 and f_2.

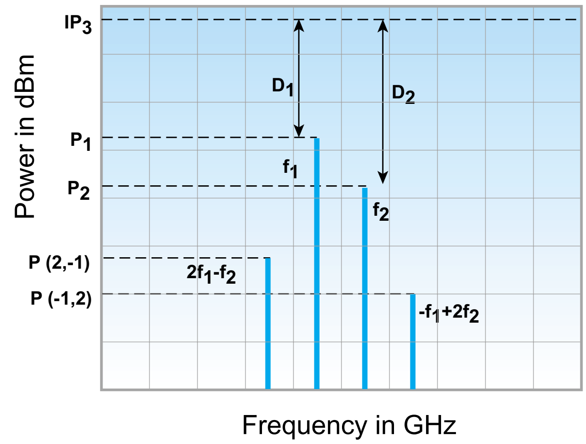

Figure 12: Frequency Spectrum for Unequal Signals

Harmonic Distortion:

Harmonic distortion is also a result of non-linearity in the amplifier transfer function and manifests itself as output signals at integral multiples of the input signal frequency. Since harmonic distortion is a function of input power, it is usually specified in terms of the relative level of the harmonics with respect to the power of the fundamental signal.

Typical second harmonic outputs are –20 dBc to –25 dBc at the amplifier 1 dB compression point. Third harmonic outputs are generally 10 dB less than second harmonics.

It should be noted that in very broadband amplifier designs (amplifier design may be broadband although the specified operating bandwidth is narrowband), significant gain may be present at harmonic frequencies thereby magnifying the harmonic output problem. Conversely, if the amplifier has significant attenuation at the harmonic frequency, harmonic output may be much lower than the numbers given above.

Non-Harmonic Spurious:

These spurious signals are not related to the harmonics or intermodulation products of the fundamental input signals, In general, non-harmonic spurious signals result from poor performance of other system components, faulty amplifier design, or insufficient user specifications relating to the amplifier. The following discussion describes the most common problems that can result in spurious signals.

• Amplifier Instability

Amplifiers that are normally stable and spurious free when terminated in well matched loads at the input and output (normal test condition) may become unstable when terminated in high VSWR loads and/or when signals are present which drive the amplifier into non-linear operation.

Spurious oscillations of this type are often at frequencies well out of the operating band of the amplifier. They are most commonly at very low frequencies that may be several orders of magnitude below the specified operating band. In this case, the spurious signals may not be noticeable on a spectrum analyzer until a signal is present at the amplifier input. The spurii are then manifested as modulation sidebands on the output signal.

In order to prevent occurrence of this problem, Microwave dB designs are tested with infinite VSWR variable phase loads at the input and output with the simultaneous application of sufficient input power to saturate the amplifier (Figure 13). The input frequency and power are varied over a wide range as the phase of the infinite VSWR load is changed. The output is observed on a wide-band spectrum analyzer (1 MHz IF bandwidth).

Microwave dB amplifiers are generally considered to be spurious-free when no spurri can be observed under the above conditions at input powers at least 6 dB above the 1 dB compression point (most amplifiers will be spurious-free at input powers up to burnout levels).

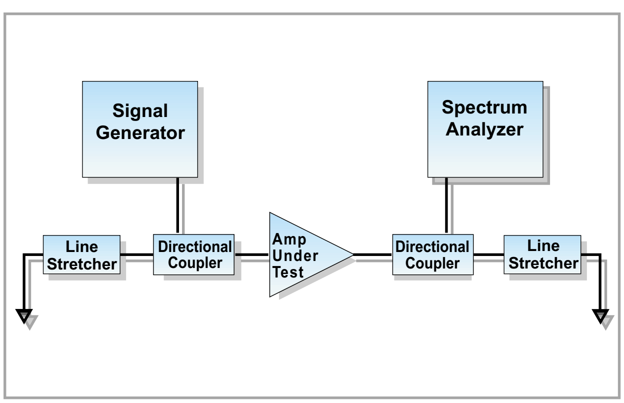

Figure 13: Stability Test Set

• Power Supply Induced Spurious Signals

The gain and phase of all amplifiers will, to some degree, vary as a function of supply voltage. If the power supply voltage contains sufficient ripple or noise components to modulate the amplifier gain transfer function, the output signal will contain amplitude and/or phase sidebands at the power supply frequencies.

The sensitivity of amplifier parameters to supply voltage ripple and noise is generally minimized through careful amplifier design as well as inclusion of voltage regulators as part of the amplifier assembly. However, the characteristics of the available power supply (other than just voltage and current) should be included in a specification to ensure compatibility of the amplifier with the overall system.

• RFI, Conducted and Radiated

Conducted RFI consists of interfering signals that appear on any of the connections to the amplifier. In general, the amplifier manufacturer is concerned (unless otherwise specified) with RFI signals on the power supply connections.

One effect of RFI on the power terminal is essentially the same as power supply ripple, i.e. the amplifier gain characteristics are modulated with resulting sidebands as described above. Another equally important aspect, particularly when the interfering signal is within the amplifier band, is the re-radiation of the interfering signal into the amplifier cavity and subsequent amplification. Both of the above problems are usually solved by careful design of voltage regulation and filtering circuits.

Radiated RFI is basically leakage of signals into the amplifier cavity and is usually a result of poor shielding.

Microwave dB amplifiers address applicable portions of the MIL specifications governing conducted and radiated RFI. Any questions regarding these important but often overlooked parameters should be addressed to the Microwave dB Applications Engineering Staff.

VSWR

This term is an acronym for Voltage Standing Wave Ratio and is a measure of the degree of match of a load relative to the system characteristic impedance.

Return Loss

Virtually all measurements of VSWR using modern test equipment are performed by measuring the Return Loss of the test unit. This parameter is simply the square of the reflection coefficient expressed in dBs. These parameter relationships are defined in Figure 14.

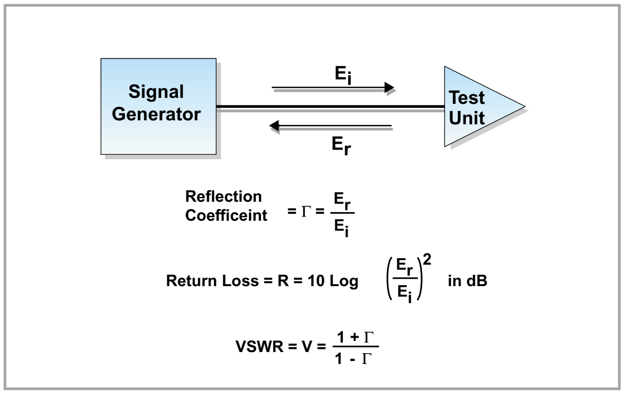

Figure 14: VSWR Relationships

Return Loss Measurement:

Figure 15 shows a typical test setup for measuring return loss. With the short circuit connected to the directional coupler (or bridge) a reference is obtained on the indicator. The test unit is then connected and the return loss is obtained from the difference in (dB) between the two measurements. From the previous equations, the return loss, R, is related to VSWR as:

$$\text{VSWR} = \frac{1 + 10^{-0.05R}}{1 - 10^{-0.05R}}$$

The accompanying table presents VSWR and reflection coefficients for values of return loss, R, from 1 dB to 30 dB. It should be noted that the calculations for the table assume infinite directivity. In practice, the effective directivity of the VSWR Test Set is the vector sum of coupler directivity, D_C, and the reflections, R_M, at or before the test port:

$$\text{Effective Directivity} =$$

$$D_E = 20\,\text{Log}_{10}\,(10^{-0.05D_c} + 10^{-0.05R_m})$$

Thus, a directional coupler with an inherent directivity of D_C = 20dB followed by a transition (e.g. type "N" to type "SMA") with return loss, R_M = 20dB (VSWR = 1.22:1) will have an effective directivity of D_E = 14dB. For purposes of VSWR measurement accuracy, the effective directivity, D_E, should be at least 10dB better than the return loss specification.
In cases where small values of return loss are being measured, extreme care must be exercised to minimize measurement uncertainty caused by effective directivity. The error associated with the effective directivity may be expressed by the following approximation providing $D_E \geq 20$dB:

$$\text{Measured VSWR} \approx V_{actual}\left(\frac{1 \pm 2\,\text{x}10^{-0.05D_E}}{1 - 10^{-0.1R}}\right)$$

Thus, with an actual VSWR of 1.5, R = 14 dB, and an effective directivity of 20 dB, the measured VSWR could be 1.701:1 or 1.299:1 depending on the phase relationships of the test components. The effective directivity may vary from test set to test set and is frequently the source of data correlation problems.

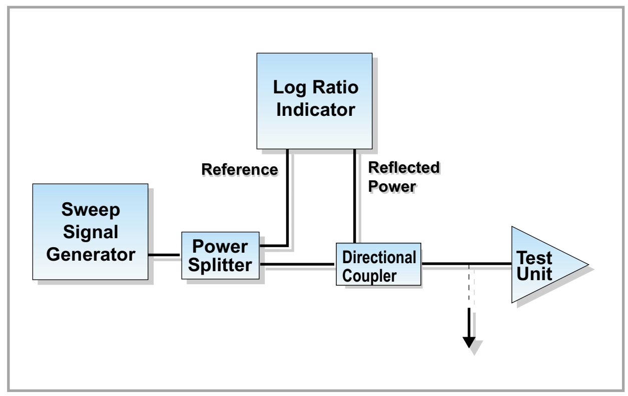

Figure 15: Return Loss Test Set

Return Loss Chart

Return Loss dB	VSWR	Reflection Coefficient	Loss dB	VSWR	Reflection Coefficient
1	17.3910	0.8913	16	1.3767	0.1585
2	8.7242	0.7943	17	1.3290	0.1413
3	5.8480	0.7079	18	1.2880	0.1259
4	4.4194	0.6310	19	1.2528	0.1122
5	3.5698	0.5623	20	1.2222	0.1000
6	3.0095	0.5012	21	1.1957	0.0891
7	2.6146	0.4467	22	1.1726	0.0794
8	2.3229	0.3981	23	1.1524	0.0708
9	2.0999	0.3548	24	1.1347	0.0631
10	1.9250	0.3162	25	1.1192	0.0562
11	1.7849	0.2818	26	1.1055	0.0501
12	1.6709	0.2512	27	1.0935	0.0447
13	1.5769	0.2239	28	1.0829	0.0398
14	1.4985	0.1995	29	1.0736	0.0355
15	1.4326	0.1778	30	1.0630	0.0316